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⑫ **EUROPEAN PATENT APPLICATION**

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㊾ Coded modulation, with unequal error protection, for fading channels.

㊲ Information is transmitted in digital form over fading channels using DPSK coded modulation incorporating multi-level coding (120) in order to provide unequal error protection for different classes of data (on K1, K2, K3) such as generated by CELP or other speech encoders (104).

FIG. 1

EP 0 544 463 A2

Background of the Invention

The present invention relates to the transmission of information in digital form over fading channels.

The increasing prominence of wireless telecommunications--particularly in the realm of digital cellular mobile radio--has given rise to the demand for improvements in the bandwidth efficiency of such systems. To this end, efficient coded modulation schemes, such as the built-in time-diversity technique disclosed in U.S. Patent 5,029,185, issued to L.F. Wei on July 2, 1991 and entitled "Coded Modulation for Mobile Radio," have been developed. Increased bandwidth efficiency is also achieved in these systems via the use of low-bit-rate speech coders, such as the so-called code-excited linear predictive (CELP) coders, which operate in the range of about 4 to 8 kilobits per second (kbps).

The low-bit-rate coders represent speech information in such a way that certain aspects of the coded information is of significantly greater importance than other aspects in terms of being able to recover intelligible speech at the receiver. In CELP coders, for example, that so-called "important information" may comprise a) the linear predictive coding (LPC) parameters, b) the pitch, and c) one information bit which indicates whether the speech was voiced or unvoiced. It is thus desirable that the transmission scheme be able to communicate the "important information" with a high degree of reliability, even at the expense--if channel conditions make it necessary--of the the other, "less important" information. Such transmission schemes are referred to herein generically as schemes which provide "unequal error protection." And it should also be noted at this point that, in general, there can be any desired number of classes of information, of varying importance, rather than being limited to, for example, two classes.

Transmission schemes which provide unequal error protection are, in fact, known in the prior art. Such known technology is exemplified by, for example, in Carl-Erik Sundberg, "Optimum Weighted PCM for Speech Signals," IEEE Transactions on Communications, Vol. COM-26, No. 6, June 1978, pp. 872-881; C.-E.W. Sundberg et al., "Logarithmic PCM weighted QAM transmission over Gaussian and Rayleigh fading channels," IEE Proceedings, Vol. 134, Pt. F, No. 6, October 1987, pp. 557-570. Wireless telecommunications of the type that the present invention is concerned with are typically carried out over so-called fading channels, by which is meant so-called Rayleigh or near-Rayleigh channels, where multiplicative Rayleigh noise is the predominant noise phenomenon. However, the prior art (for coded modulation) has generally addressed the unequal error protection problem in the context of transmission of the information over channels where additive white Gaussian noise is the predominant noise phenomen-

on, and those schemes will not perform effectively if used for the fading environment. It can also be noted that schemes which provide unequal error protection in conjunction with binary and quaternary phase shift keyed signals are known in the prior art. See "Rate-compatible punctured convolutional codes for digital mobile radio," by J. Hagenauer, N. Seshadri and C.-E.W. Sundberg in IEEE Transactions on Communications, Vol. 37, No. 7, July 1990. However, these schemes are limited to a maximum of two bits per symbol, irrespective of the signal-to-noise ratio (SNR) of the channel and are thus bandwidth-inefficient. Moreover, even when achieving two bits per symbol, these schemes are power-inefficient.

There thus remains in the art the need for effective, bandwidth- and power-efficient transmission schemes which can provide unequal error protection in fading channel environments.

Summary of the Invention

The present invention meets that need. In particular, in accordance with the principles of the invention, the information is coded using a multi-level channel code, by which is meant that a) each class of information is redundancy coded using a different, respective channel code and that b) the resulting multi-level-coded words select for transmission signal points of a predetermined signal constellation. In preferred embodiments, the minimum Hamming distance (defined below) for the code used for any particular class of information is grater than the minimum Hamming distance for the code used for any less important class of information. The error probability for channels characterized by Rayleigh fading assuming adequate interleaving (as described below), and coherent or differential coherent detection, at high SNR is, to a first, and typically adequate, approximation, proportional to the reciprocal of the signal-to-noise ratio (SNR) raised to the power of the minimum Hamming distance (assuming the use of, for example, a maximum likelihood decoding strategy). That is, for the ith class of data, class i,

$$P_e(i) \approx \gamma_i [1/SNR]^{d_{Hi}},$$

where d_{Hi} is the minimum Hamming distance, γ_i is a proportionality constant and $P_e(i)$ is the probability of error upon decoding. Different levels of error protection are thus provided for the different classes of data.

As is explained in detail hereinbelow, the proportionality constant γ_i depend primarily on the ratio of a) the average number of nearest neighbors for the ith class of data to b) the product distance for that class. These parameters are determined by a) the channel codes and b) the signal constellation design, the latter including, for example, the constellation geometry, the labelling of the signal points and the manner in which the outputs of the channel codes map into

the labels. Thus, in accordance with a feature of the invention, attainment of particular desired error probabilities for the vinous different data classes can be facilitated by appropriate joint selection of channel codes and signal constellation design.

The prior art does disclose the use of multi-level codes in a fading channel environment, but only in the context of providing equal error protection. Exemplary is N. Seshadri and C.-E.W. Sundberg, "Multi-level codes with large time-diversity for the Rayleigh fading channel," Conference on Information Sciences and Systems, Princeton, N.J., pp. 853-857, March 1990.

Brief Description of the Drawing

FIG. 1 is a block diagram of a transmitter embodying the principles of the invention;
FIG. 2 is a block diagram of a receiver for signals transmitted by the transmitter of FIG. 1;
FIG. 3 depicts a signal constellation useful in explaining the principles of the invention;
FIG. 4 is a block diagram of a multi-level encoder used in the transmitter of FIG. 1; and
FIGS. 5-7 depict signal constellations that can illustratively be used by the transmitter of FIG. 1.

Detailed Description

Before proceeding with a description of the illustrative embodiments, it should be noted that various of the digital signaling concepts described herein are all well known in, for example, the digital radio and voiceband data transmission (modem) arts and thus need not be described in detail herein. These include such concepts as multidimensional signaling using 2N-dimensional channel symbol constellations, where N is some integer, trellis coding; scrambling; passband shaping; equalization; Viterbi, or maximum-likelihood, decoding; etc. These concepts are described in such U.S. patents as U.S. 3,810,021, issued May 7, 1974 to I.Kalet et et al; U.S. 4,015,222, issued March 29, 1977 to J. Werner, U.S. 4,170,764, issued October 9, 1979 to J. Salz et al; U.S. 4,247,940, issued January 27, 1981 to K. H. Mueller et al; U.S. 4,304,962, issued December 8, 1981 to R. D. Fracassi et al; U.S. 4,457,004, issued June 26, 1984 to A. Gersho et al; U.S. 4,489,418, issued December 18, 1984 to J. E. Mazo; U.S. 4,520,490, issued May 28, 1985 to L. Wei; and U.S. 4,597,090, issued June 24, 1986 to G. D. Forney, Jr.

It may also be noted before proceeding that various signal leads shown in the FIGS. may carry analog signals, serial bits or parallel bits, as will be clear from the context.

Turning now to FIG. 1, which illustratively depicts a digital cellular mobile radio terminal such as might be installed in an automobile, speech signal source 101 generates an analog speech signal representing speech information or "intelligence," which is passed on to speech encoder 104, illustratively a CELP coder of the type described above. The latter generates a digital signal comprised of a stream of data, or data elements, in which at least one subset of the data elements represents a portion of the information, or intelligence, that is more important than the portion of the information, or intelligence, represented by the rest of the data elements. Illustratively, each data element is a data bit, with an average $K = k_1 + k_2 + k_3$ information bits being generated for each of a succession of M symbol intervals. The symbol intervals are comprised of N signaling intervals, where 2N is the number of dimensions of the constellation (as described below). The signaling intervals have duration of T second and, accordingly, the symbol intervals each have a duration of NT seconds. The embodiments explicitly disclosed herein happen to use two-dimensional constellations, i.e., N = 1. For those embodiments, then, the signaling intervals and symbol intervals are the same.

Of the aforementioned K information bits, the bits within the stream of k_1 bits per M symbol intervals, appearing on lead 105, are more important than the k_2 bits per M symbol intervals, appearing on lead 106, and in turn these bits are more important than the k_3 bits per M symbol intervals appearing on lead 107. The bits on these three leads are referred to herein as the class 1, class 2 and class 3 bits, respectively.

The bits on leads 105-107 are independently scrambled in scramblers 110-112, which respectively output on leads 115-117 the same number of bits per M symbol intervals as appear on leads 105-107, respectively. (In particular specific embodiments, scrambling may not be required.) Scrambling is customarily carried out on a serial bit stream. Thus although not explicitly shown in FIG. 1, scramblers 110-112 may be assumed to perform a parallel-to-serial conversion on their respective input bits prior to scrambling and a serial-to-parallel conversion at their outputs. The signal is then channel encoded and mapped. In particular, the bits on leads 115-117 are extended to multi-level coder 120. As described in detail hereinbelow, the latter illustratively includes three channel encoders which respectively receive the bits on leads 115-117. These encoders generate, for each block of M symbol intervals, a block of M multibit (illustratively 3-bit), multi-level-coded, words. M is greater than $(k_1 + k_2 + k_3)/3$ so that the multi-level encoder is a redundancy coder, i.e., more bits are output than are input. The multi-level code implemented by encoder 120 to generate the multi-level-coded words on lead 121 has so-called built-in time-diversity. Those bits comprise a multi-level-coded word and their values jointly select, or identify, a particular channel symbol of a predetermined constellation of channel symbols (such as the constellation of FIG. 5

as described in detail hereinbelow). Complex plane coordinates of the identified channel symbols are output by constellation mapper 131, illustratively realized as a lookup table or as a straightforward combination of logic elements. The complex channel symbols are interleaved in standard fashion by an interleaver 141 in order to be able to take advantage of the built-in time-diversity of the multi-level code. Since such interleaving is standard, it suffices for present purposes simply to note that the function of the interleaver is to provide adequate time-separation between signal points which comprise a coded block to ensure that fading events associated with those signal points are, in general, independent. The output of interleaver 141 is applied to modulator 151 and the resultant analog signal is then broadcast via antenna 152 over a free space communication channel to a remote digital cellular mobile radio cell site.

In order to understand the theoretical underpinnings of the invention, it will be useful at this point to digress to a consideration of FIG. 3. The latter depicts a standard two-dimensional data transmission constellation of the general type conventionally used in digital cellular mobile radio. In this standard scheme--conventionally referred to as differential phase shift keyed (DPSK) modulation--data words each comprised of two information bits are each mapped into one of four possible two-dimensional channel symbols. The phase angle of each signal point (measured from the positive X axis) indicates a change that the phase of the transmitted signal must undergo in order to transmit the bit pattern associated with that particular signal point. More particularly, so-called $\pi/4$-shifted DPSK, in which the entire constellation is rotated by 45 degrees in successive signaling intervals, can be used. (Non-differential (coherent) PSK, where each signal point represents the associated bit pattern directly, could alternatively be used in appropriate applications.) Each channel symbol has an in-phase, or I, coordinate on the horizontal axis and has a quadrature-phase, or Q, coordinate on the vertical axis. The signal points have equal amplitude--lying on a circle of radius 1--so that, on each axis, the channel symbol coordinates are $\pm\frac{1}{\sqrt{2}}$. Thus the distance between each symbol and each of the symbols that are nearest to it is the same for all symbols--that distance being $\sqrt{2}$. As a result of this uniform spacing, essentially the same amount of noise immunity is provided for both information bits.

We now define some useful terminology: "Hamming distance," "minimum Hamming distance," "number of nearest neighbors," and "product distance." In particular, the "Hamming distance" between any two distinct sequences of signal points selected from the constellation is the number of positions within the two sequences at which the signal points differ. The "minimum Hamming distance" is the minimum over all

such Hamming distances, i.e., taking into account all possible pairs of sequences. Since we are assuming at this point an uncoded transmission scheme, all possible sequences of signal points are allowed. This being so, the minimum Hamming distance for an uncoded transmission scheme is "1". A code which has so-called built-in time-diversity has a minimum Hamming distance which is greater than "1", the measure of that time-diversity being, in fact, the minimum Hamming distance. The "product distance" is the product of all non-zero squared Euclidean distances between the corresponding signal points of the sequences that are at minimum Hamming distance from each other--known as the "nearest neighbor sequences." For the present case, the product distance is "2". The "number of nearest neighbors" is the average number of sequences that are at the minimum Hamming distance to any transmitted sequence with a product distance of 2 and, in this case, the "number of nearest neighbors" is "2".

As is well known, it is possible to provide improved noise immunity without sacrificing bandwidth efficiency (information bits per signaling interval) using a coded modulation approach in which an "expanded" two-dimensional constellation having more than (in this example) four symbols is used in conjunction with a trellis or other channel code. For example, the above-cited U.S. patent 5,029,185 discloses the use of expanded PSK constellations in combination with trellis and block codes to provide enhanced noise immunity of more than 10 dB over the uncoded case of FIG. 3 while providing approximately two bits per signaling interval.

In the case of FIG. 3 hereof, the same amount of noise immunity is provided for all information bits. Thus, as noted above, there remains in the art the need for effective, bandwidth- and power-efficient transmission schemes which can provide unequal error protection in fading channel environments--a need that is met by the present invention.

In particular, in accordance with the principles of the invention, the information is coded using a multi-level channel code, by which is meant that a) each class of information is channel coded using a different, respective channel code and that b) the resulting coded outputs select for transmission signal points of a predetermined signal constellation.

To this end, and referring to FIG. 4, there is shown an illustrative embodiment of multi-level--illustratively three-level--channel coder 120. As noted earlier, the three classes of data are received on leads 115-117, from most- to least-important. The bit streams for these three classes are denoted i_1, i_2 and i_3 and are applied to respective channel encoders 40i, $i = 1, 2, 3$, i.e., encoders 401, 402 and 403, which respectively implement redundancy codes C_i, $i = 1, 2, 3$, i.e., codes C_1, C_2 and C_3. The output bit streams from the three channel encoders--which are buffered

in buffers 411, 412 and 413, as described below--are denoted $\underline{b}_1$, $\underline{b}_2$ and $\underline{b}_3$ and appear on leads 122-124, respectively. The constituent elements of the output bit streams can be represented as follows, in which the superscripts denote time:

$$\underline{b}_3 = (b_3^1, b_3^2, ..., b_3^M)$$
$$\underline{b}_2 = (b_2^1, b_2^2, ..., b_2^M)$$
$$\text{and } \underline{b}_1 = (b_1^1, b_1^2, ..., b_1^M) .$$

The mulit-level encoder output is

$$\underline{b} = (b^1, b^2, ..., b^M)$$
$$\text{where } b^j = 4b_3^j + 2b_2^j + b_1^j .$$

The index b^j then constitutes an address applied to constellation mapper 131 to select a particular signal point of an illustrative 8-PSK constellation which is shown in FIG. 5. Note, in particular, that each of the eight signal points of the constellation has an associated label--shown in both binary and, in parentheses, decimal form--which is used as the address. Various ways of assigning the signal point labels to achieve certain overall error probabilities for particular codes and particular constellations are discussed hereinbelow.

The codes respectively implemented by encoders 40i, i = 1, 2, 3, are codes which are characterized by the parameter set (M, k_i and d_{Hi}), i = 1, 2, 3. Here, M, introduced above, is the block length of the code; k_i, introduced above, is the number of information bits required to be applied to encoder 40i to generate the output block; and d_{Hi} is the minimum Hamming distance (defined above) of code C_i. The three codes are illustratively as follows: Code C_1 is a (4,1,4) binary repetition code consisting of codewords 0000 and 1111; code C_2 is a (4,3,2) binary parity check code with even parity; and code C_3 is a (4,4,1) code, which means that, in this example, no redundancy is added. That is, class 3 is not coded at all. Since M = 4, and since each signal point is a two-dimensional signal point, the overall coded modulation scheme is 8-dimensional (8D). The class 1 bits--which are the most important--are encoded by code C_1; the class 2 bits--which are the second-most important--are encoded by code C_2; the class 3 bits--which are the third-most important--are encoded by code C_3. In order to provide three-bit addresses to constellation mapper 131 on an ongoing, regular basis, buffers 411, 412 and 413, each of length M = 4, are provided, as already noted, to buffer the outputs of encoders 401, 402 and 403. It is thus seen that K = $k_1 + k_2 + k_3$ = 8. That is, 8 information bits are transmitted for each block of M = 4 symbol intervals, yielding a bit rate of 2 bits per signaling interval. Note that 1/8 = 12.5 % of the data is in class 1; 3/8 = 37.5 % of the data is in class 2; and 4/8 = 50.0 % of the data is in class 3.

In order to advantageously use the minimum Hamming distance separation, i.e., the built-in time-diversity of a code, it is necessary that symbols within any one block of M signal points be subject to inde-

pendent fading. In practice this is achieved by way of the interleaving provided by interleaver 141 as described above.

We consider, now, specifically the error probability for each of the data classes in this example. We first note that the typical cellular mobile radio channel is a Rayleigh fading channel. The probability of error for such channels is, to a first, and typically adequate, approximation, proportional to the reciprocal of the signal-to-noise ratio (SNR) raised to the power of the minimum Hamming distance (assuming the use of, for example, a maximum likelihood decoding strategy). That is, for the ith class of data

$$P_e(i) \approx \gamma_i (\frac{1}{SNR})^{d_{Hi}},$$

where

    d_{Hi} is the minimum Hamming distance

    P_e (i) is the probability of error upon decoding

    and γ_i is a proportionality constant .

(Hereinafter, for convenience, we will use "P_e (i) = "rather than "P_e (i) $\approx$".) Therefore, since the minimum Hamming distance for class 1 is "4," which is greater than that for any less important class of data for which the minimum Hamming distance is "2", the bit error probability for the former is, in accordance with the invention, better (i.e., lower) than that for the latter, assuming some minimum SNR and ignoring, in the first instance, the contributions of the proportionality constants γ_i.

Of course, it is readily seen that the proportionality constants γ_i do also make a contribution to $P_e(i)$. Each γ_i, in particular, is is the ratio of a) the average number of nearest neighbors for the ith class of data to b) the so-called product distance for that class. (Hereinafter, for convenience, we will use "γ_i =" rather than "γ_i $\approx$".) These parameters are determined by codes chosen, as well as the signal constellation design, which includes, for example, the constellation geometry, the labelling of the signal points and the manner in which the outputs of the channel codes map into the labels. Thus, in accordance with a feature of the invention, attainment of particular desired error probabilities for the various different data classes can be facilitated by appropriate joint selection of channel codes and signal constellation design. In virtually all instances of practical interest, the code selection and constellation design will be such that the values of the γ_i will not change the result that $P_e(i) < P_e(2) < P_e(3)...$ for $d_{Hi} > d_{H2} > d_{H3}....$

The product distance for the class 1 data is $0.587^4 = 0.119$, which can be verified by noting that the minimum Hamming distance of code C_1 is "4" and that the squared Euclidean distance between nearest neighbors of the signal constellation which differ in the bit that is addressed by the output of code C_1--bit b_1--is 0.587. The number of nearest neighbors for this code is 8 since, corresponding to any transmitted multi-level-coded words, there are 8 other multi-level-

coded words which are at a Hamming distance of 4 with a product distance of 0.119. Thus, $\gamma_i = 8/0.119 = 67.23$.

For class 2 data, the product distance is $2^2 = 4$, which can be verified by noting the minimum Hamming distance of code C_2 is "2" and that the squared Euclidean distance between nearest neighbors of the signal constellation which differ in the bit that is addressed by the output of this code is 2.0. The number of nearest neighbors for this code is 6 since for each multi-level-coded word (assuming i_1 has been decoded correctly), there are 6 multi-level-coded words that are at a Hamming distance of 2 that can result in an error in information sequence i_2. Thus, $\gamma_2 = 6/4 = 1.5$.

For class 3 data, the product distance is 2.0, which can be verified by noting the minimum Hamming distance for an uncoded case--which is what code C_3 is in this example--is "1" since the squared Euclidean distance between nearest neighbors of the signal constellation which differ in the bit that is addressed by the output of this code is 2.0. The number of nearest neighbors for this code is "1" since there is only one neighbor at squared Euclidean distance "2.0" (assuming that the bits encoded by codes C_2 and C_1 have been decoded correctly). Thus, $\gamma_3 = \frac{1}{2} = 0.5$.

Overall, then,
$$\gamma_1 = 8/0.119 = 67.2,$$
$$\gamma_2 = 6/4 = 1.5,$$
$$\gamma_3 = 1.0/2.0 = 0.5,$$
so that data classes 1, 2 and 3 have the following error probabilities:
$$P_e(1) = 67.2[1/SNR]^4,$$
$$P_e(2) = 1.5[1/SNR]^2,$$
$$P_e(3) = 0.15[1/SNR]^1.$$
Note that as desired--at least when the SNR is greater than some minimum--the greatest level of error protection (lowest error probability) is provided to class 1; the second-greatest level of error protection is provided to class 2; and the third-greatest level of error protection is provided to class 3.

(In this example, as well as all the other examples given herein, it is assumed that the interleaver provides sufficient separation in time between the signal points within a block of M signal points to ensure that fading events associated with those signal points are, in general, independent.)

It will be immediately apparent, then, that using different codes--to provide different minimum Hamming distances--and different constellations--to provide different proportionality constants--various different sets of level of protection can be provided. A few different possibilities--which are, of course, illustrative and not in any sense exhaustive--will now be presented. We begin by changing the codes while continuing to use the constellation and signal point labelling of FIG. 5.

Consider, for example, a 16-D block coded modulation scheme which uses the following set of codes:

C_1 is an (8,4,4) extended Hamming code;

C_2 is an (8,7,2) binary pity check code with even pity;

C_3 is an (8,7,2) binary parity check code with even parity.

It is thus seen that $K = k_1 + k_2 + k_3 = 18$. That is, 18 information bits are transmitted for each block of M = 8 symbol intervals, yielding a bit rate of 2.25 bits per signaling interval. Note that 4/18 = 22.2 % of the data is in class 1; 7/18 = 39.9% of the data is in class 2; and 7/18 = 39.9% of the data is in class 3.

Overall, then,
$$\gamma_1 = 224/0.119 = 1882.35,$$
$$\gamma_2 = 28/4.0 = 7.0,$$
$$\gamma_3 = 28/4.0 = 7.0,$$
so that data classes 1, 2 and 3 have the following error probabilities:
$$P_e(1) = 1882.35[1/SNR]^4,$$
$$P_e(2) = 7[1/SNR]^2,$$
$$P_e(3) = 7[1/SNR]^2.$$

It will be readily seen that, in this case, the same level of error protection is provided to classes 2 and 3, yielding, in effect, a two- rather than a three-level unequal error protection scheme. However, as will be seen later, using a different constellation design--while not affecting the minimum Hamming distance-can result in a change of the proportionality constants, thereby providing different levels of error protection for classes 2 and 3.

Now we consider a 32-D block coded modulation scheme which uses the following set of codes:

C_1 is an (16,11,4) extended Hamming code;

C_2 is an (16,15,2) binary parity check code with even parity;

C_3 is an (16,15,2) binary parity check code with even parity.

It is thus seen that $K = k_1 + k_2 + k_3 = 41$. That is, 41 information bits are transmitted for each block of M = 16 symbol intervals, yielding a bit rate of 2.56 bits per signaling interval. Note that 11/41 = 26.8 % of the data is in class 1; 15/41 = 36.6% of the data is in class 2, and 15/41 = 36.6% of the data is in class 3.

Overall, then,
$$\gamma_1 = 2240/0.119 = 18823.5,$$
$$\gamma_2 = 120/4 = 30,$$
$$\gamma_3 = 120/4 = 30,$$
so that data classes 1, 2 and 3 have the following error probabilities:
$$P_e(1) = 18823.5[1/SNR]^4,$$
$$P_e(2) = 30[1/SNR]^2,$$
$$P_e(3) = 30[1/SNR]^2.$$

Note that in this case, the error probabilities are approximately the same as in the previous example. However, the bits per symbol are different.

Now we consider a coded modulation scheme af-

fording two, rather than three, levels of error protection and in which a binary convolutional code is used for protecting the class 1 data and a block code is used for protecting the class 2 data. The codes, in particular, are:

C₁ is a rate $R = 1/2$ convolutional code;

C_2 is a $(L,L-1,2)$ binary parity check code with even parity.

Code C_1, more particularly, is a maximum free distance (d_{free}) code for any memory of the code that is desirable and is of the type described, for example, in J.G. Proakis, Digital Communications, 2nd Ed., McGraw-Hill, 1989. The two bits output by code C_1 determine bits b_1 and b_2 and the output from code C_2 determines b_3. In this case, the total number of bits transmitted in a block of length L is $K = (2L - 1)$, yielding a bit rate of 2-(1/L) bits per signaling interval. Approximately 50% of the bits are in each of the classes 1 and 2, for reasonably large L. For example, if code C_1 is a memory 2 convolutional code, and $L = 10$, then

$$\gamma_1 = 1/2.25 = 0.44,$$
$$\gamma_2 = 45/16 = 2.81.$$

Moreover, as is well known, the parameter for convolutional codes which corresponds to the role of the minimum Hamming distance for block codes is, in this case, $(d_{free}-2) = 3$. In general, an upper bound on the value of time diversity for convolutional codes used in the manner described here is $d_{free}-2$. For memory 2 code, this upper bound is achieved. Overall, then, data classes 1 and 2 have the following error probabilities:

$$P_e(1) = 1.0/2.25 [1/SNR]^3,$$
$$P_e(2) = 45/16 [1/SNR]^2.$$

(For this example, in order to achieve the above error probabilities, and, in particular, the product distances indicated, the labeling assignment shown in FIG. 5 should be changed. The 000 label is assigned to the same signal point of the constellation. The other labels, reading counter-clockwise, are then, for this case, 001, 011, 010, 100, 101, 111 and 110.)

The foregoing examples thus illustrate that by using codes having various a) minimum Hamming distances (or convolutional code free distance) and b) block code lengths, one can obtain various levels of error protection, overall bits per signaling interval, and/or the fraction of bits allocated to the various classes. As noted earlier, yet further flexibility is provided by using various different constellations, thereby changing the proportionality constants γ_i.

FIGS. 6 and 7 show respective 8-point PSK constellations in which the signal points are non-uniformly spaced and for which, as a result one can provide a) numbers of nearest neighbors and/or b) product distances that are different from those obtained for the constellation of FIG. 5. In this way, different proportionality constants γ_i can, advantageously, be obtained.

Consider the use of these constellations in conjunction with the first set of codes described above for the 8-dimensional block code.

In the constellation of FIG. 6, in particular, the squared minimum Euclidean distance between any two signal points that differ in bit b_1--that is the bit that is output by encoder 401--is 2.0. Hence using this constellation in place of the constellation of FIG. 5 increases the product distance from 0.119 to 16 for class 1. However, the product distance for class 2 decrease from 4 to 0.43, while the product distance for class 3 decrease from 2.0 to 1.0. The number of nearest neighbors for class 1 decrease from 8 to 1, while for classes 2 and 3 it remains the same as it was, at 6 and 1, respectively. The proportionality constants thus have the values

$$\gamma_2 = 1/16 = 0.0625,$$
$$\gamma_2 = 6/0.43 = 13.95,$$
$$\gamma_3 = 1.0/0.43 = 2.33,$$

Note, then, that the overall result is to provide even greater error protection (lower error probability) for class 1 at the expense of less error protection for classes 2 and 3. Thus, the minimum SNR at which class 1 data bits are subject to a better error probability is strictly lower with the constellation of FIG. 6 than with that of FIG. 5.

In the constellation of FIG. 7, in particular, the squared minimum Euclidean distance between any two signal points that differ in either of bits b_1 or b_2--that is the bits that are output by encoders 401 and 402--is 1.0. Hence using this constellation in place of the constellation of FIG. 5 increases the product distance from 0.119 to 1.0 for class 1. However, the product distance for class 2 decreases from 4.0 to 1.0, while the product distance for class 3 decreases from 4 to 0.43. The number of nearest neighbors for class 1 decreases from 8 to 1; for class 2, it remains at 6; and for class 3, it remains the same as it was at 1. The proportionality constants thus have the values

$$\gamma_1 = 1/1 = 1,$$
$$\gamma_2 = 6/1 = 6,$$
$$\gamma_3 = 1/0.43 = 2.32.$$

Here, the overall result is to provide a greater separation, in terms of level of error protection between classes 2 and 3 than with either of the other two constellations. Class 1 still has more protection than when the constellation of FIG. 5 is used, but not as much as when the constellation of FIG. 6 is used.

It may also be possible to use codes which have the same minimum Hamming distance, in which case unequal error protection can nonetheless be obtained by having different values of γ_i via, for example, the use of non-uniform signal constellations such as those of FIGS. 6 and 7.

We turn, now, to the receiver of FIG. 2.

In particular, the analog cellular mobile radio signal broadcast from antenna 152 is received by antenna 201 and is thereupon subjected to conventional front-end processing, which includes at a minimum,

demodulation and A/D conversion. Demodulation can be carried out by any of the known techniques, such as coherent demodulation, differentially coherent demodulation, non-coherent demodulation, etc. The front-end processing may also include such other processing as equalization, timing recovery, automatic gain control, etc., as is well known in the art. The output of front-end processing 211 is applied to de-interleaver 221, which perform the inverse task of interleaver 141 in the transmitter, thereby restoring the signal points to their original order. The de-interleaver output is passed on to multi-level decoder 231, which performs the task of recovering the information bits that were encoded by multi-level encoder 120. In the general case, a maximum-likelihood decoding algorithm is used. In particular, if the multi-level code is sufficiently simple, an exhaustive table-lookup approach can be used. For more complex codes, then the Viterbi algorithm could be used if the constituent codes of the multi-level code allow for a finite-state decoder realization. If the number of states of the multi-level code is too great to permit a practical implementation of a maximum-likelihood decoder, then multi-stage decoding can be used, this approach being described, for example, in A. R. Calderbank, "Multilevel codes and multi-stage decoding," IEEE Transactions on Communications, Vol. 37, pp. 222-229, March 1989. Enhanced multi-stage decoding as described in N. Seshadri and C.-E. W. Sundberg, "Multi-level codes with large time-diversity for the Rayleigh fading channel," Conference on Information Sciences and Systems, Princeton, N. J., pp. 853-857, March 1990, can also be used. More specifically in that case, if any of the constituent codes are complex, e.g. Reed-Solomon, block codes, then any of the known decoding techniques that approximate the performance of the maximum-likelihood decoder for such a code can be used to deal with that code within the multi-stage decoder.

The bits output by multi-level decoder 231 are provided in three parallel streams--corresponding to the three streams on leads 115-117--to respective descramblers 241-243, which perform the inverse function to scramblers 110-112 in the transmitter. Speed decoder 253 then performs the inverse function of speech encoder 104 of the transmitter, yielding a reconstructed speech signal that is passed on to the telephone network, including, typically, a cellular mobile radio switching system with which the receiver is co-located.

A transmitter and a receiver similar to those of FIGS. 1 and 2 would, of course, be provided at the cell site and mobile (automobile) site, respectively, to support communication in the other direction of transmission.

The foregoing merely illustrates the principles of the invention. For example, the various codes and constellations, including their dimensionalities are all illustrative. Any desired codes and constellations can be used. With respect to the codes, in particular, it should be noted that a data steam which is not actually redundancy coded, as is the case C_3 the first example hereinabove, may nonetheless be said to be coded with a rate R = 1 redundancy code. That is, "no coding" can be regarded--and, for definitional purposes herein, is regarded--as being a form of coding. With the respect to the constellation, in particular, any of various uniform or non-uniform configurations which have any desired number of signal points may be used advantageously to provide different proportionality constants.

Moreover, any desired number of levels of error protection can be supported by using multiple multi-level codes to encode respective sets of data streams and time-multiplexing the signal points addressed by the various multi-level codes.

The invention can be used to advantage in conjunction with any type of source coder (such as image, facsimile) that needs unequal error protection--not just speech coders.

It will also be appreciated that although the various functional elements of the transmitter and receiver are depicted as discrete elements, the functions of those elements will typically be carried out using appropriately programmed processors, digital signal processing (DSP) chips, etc.

Thus it will be appreciated that those skilled in the art will be able to devise numerous arrangements which, although not explicitly shown or described herein, embodying the principles of the invention and thus are within its spirit and scope.

Claims

1. Transmitter apparatus for communicating a plurality of streams of data over a channel characterized by Rayleigh fading, said apparatus CHARACTERIZED BY

    means (120) for encoding said plurality of streams using respective associated redundancy codes to generate multi-level-coded words,

    means (131) for selecting signal points of a predetermined signal constellation as a function of the values of said multi-level-coded words, and

    means (141, 151) for transmitting a signal representing the selected signal points over said communication channel, said channel being such that, for an ith one of said streams,

$$P_e(i) \approx \gamma_i [1/SNR]^{d_{Hi}},$$

where

SNR is the signal-to-noise ratio of said channel,

d_{Hi} is the minimum Hamming distance for the code which encodes said i^{th} stream of data,

$P_e(i)$ is the probability of error, upon decoding, for

said i^{th} stream of data, and

γ_i is a proportionality constant for said i^{th} stream of data which is a function of said codes and said constellation,

said codes and said constellation being chosen such that the value of $P_e(i)$ decreases for decreasing values of i.

2. The transmitter apparatus of claim 1
CHARACTERIZED IN THAT
each proportionality constant γ_i is the ratio of a) the average number of nearest neighbors for the ith class of data to b) the product distance for that class.

3. The transmitter apparatus of claim 1
CHARACTERIZES IN THAT
the signal points of said constellation are of equal amplitude.

4. The invention of claim 3
CHARACTERIZED IN THAT
said signal points are non-uniformly spaced in phase.

5. A method
CHARACTERIZED BY the steps of
encoding first and second portions of a stream of input data using first and second redundancy codes, respectively, to generate multi-level-coded words each of which includes at least one data element from each encoded portion, said first code having a minimum Hamming distance that is greater than the minimum Hamming distance of said second code,
selecting signal points of a predetermined signal constellation as a function of the values of said multi-level-coded words, and
transmitting a signal representing the selected signal points over a fading communication channel,
said constellation and codes being such that, in the transmission of said signal, the probability of error for said first portion is less than a probability of error for said second portion.

6. The method of claim 5
CHARACTERIZED IN THAT
said stream of input data represents information and wherein said first and second portions of said stream of input data respectively represent more and less important aspects of said information.

7. The method of claim 5
CHARACTERIZED IN THAT
the signal points of said constellation are of equal amplitude.

8. The method of claim 5
CHARACTERIZED IN THAT
said constellation and said selecting means are such as to modulate said multi-level-coded words using differential phase shift keying.

9. The method of claim 5
CHARACTERIZED IN THAT
said first code has a degree of built-in time-diversity which is greater than the degree of built-in time-diversity of said second code.
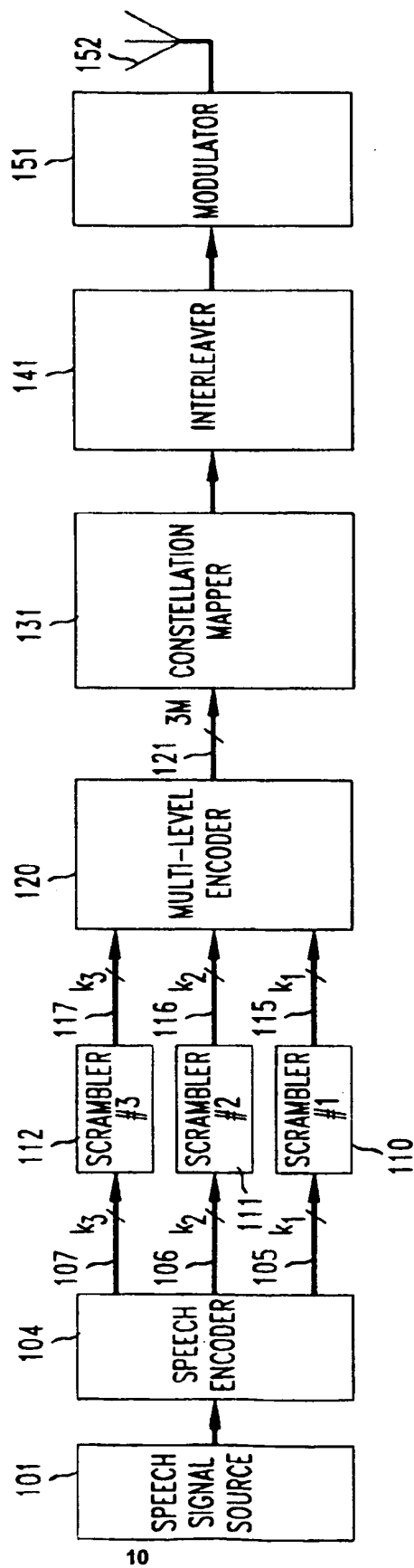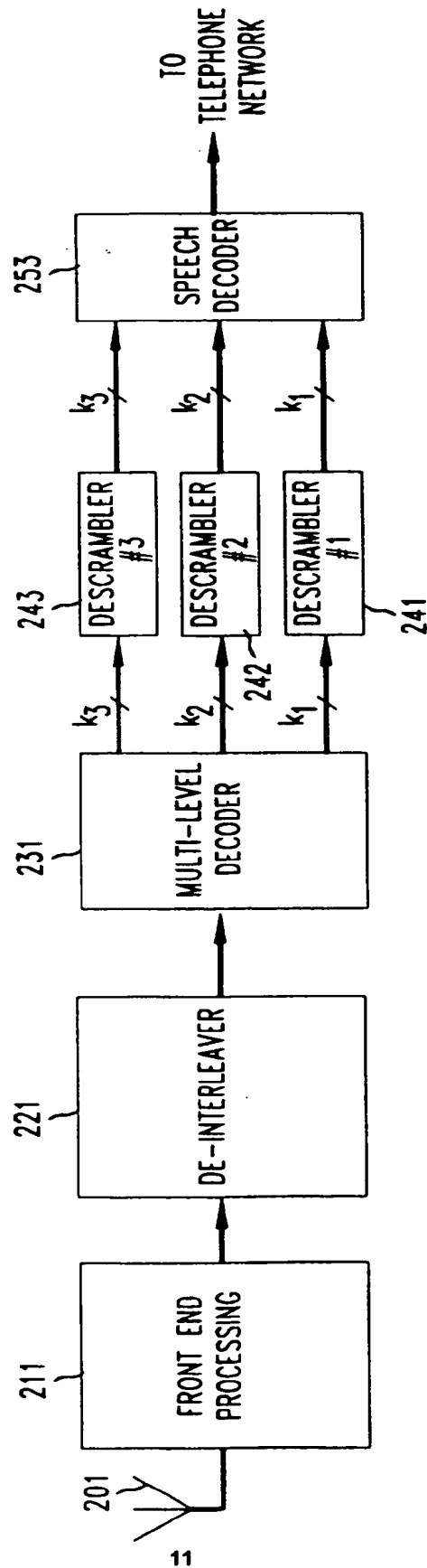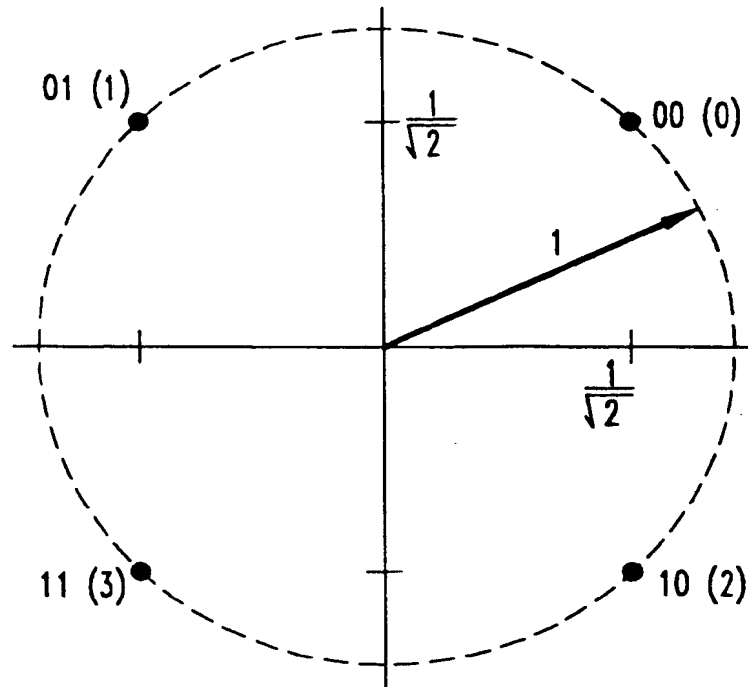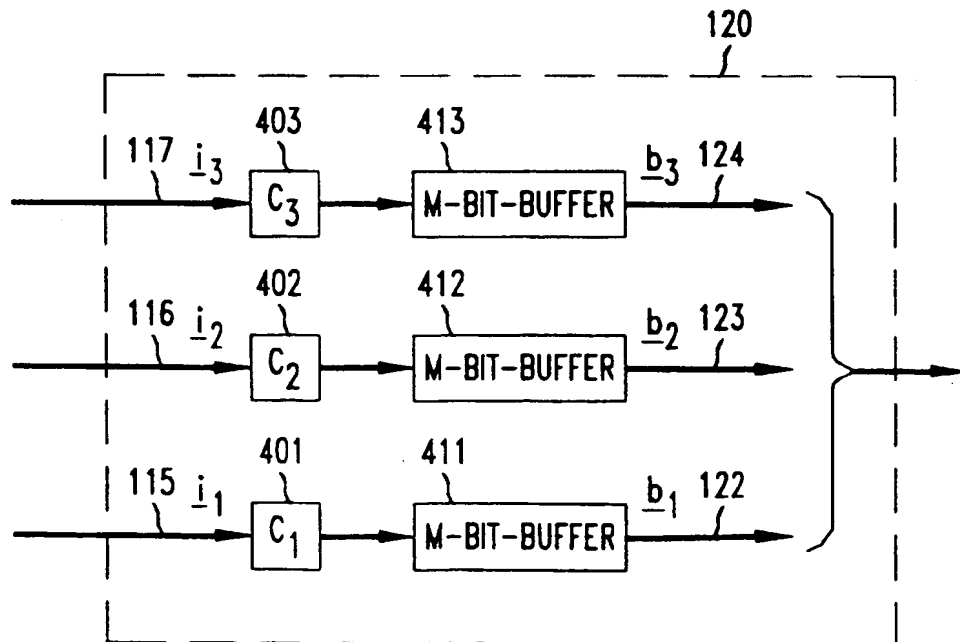
FIG. 1

FIG. 2

FIG. 3

FIG. 4

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FIG. 5

FIG. 6

FIG. 7

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�54 Coded modulation, with unequal error protection, for fading channels.

㊗ Information is transmitted in digital form over fading channels using DPSK coded modulation incorporating multi-level coding (120) in order to provide unequal error protection for different classes of data (on K1, K2, K3) such as generated by CELP or other speech encoders (104).

FIG. 1

EP 0 544 463 A3

EUROPEAN SEARCH REPORT

European Patent Office

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Category	Citation of document with indication, where appropriate, of relevant passages	Relevant to claim	CLASSIFICATION OF THE APPLICATION (Int. Cl.5)
A	CODED MODULATION AND BANDWIDTH-EFFICIENT TRANSMISSION. PROCEEDINGS OF THE FIFTH TIRRENIA INTERNATIONAL WORKSHOP ON DIGITAL COMMUNICATIONS, 8-12/9/1991 TIRRENIA, ITALY pages 305 - 316 SESHADRI N. & SUNDBERG C. E. W.: 'MULTI-LEVEL CODED MODULATIONS FOR FADING CHANNELS' * abstract * * page 307, line 17 - page 308, line 10 * * page 311, line 11 - line 15 * * page 313, line 16 - line 21 * * figure 1 *	1-9	H04L27/18
A,D	IEEE TRANSACTIONS ON COMMUNICATIONS, vol. 38, no. 7, July 1990, NEW YORK, US pages 966 - 980 HAGENAUER J., SESHADRI N. AND SUNDBERG C. E. W.: 'THE PERFORMANCE OF RATE-COMPATIBLE PUNCTURED CONVOLUTIONAL CODES FOR DIGITAL MOBILE RADIO' * abstract * * page 967, right column, line 53 - line 59 * * page 972, paragraph B *	1-9	
A	ELECTRONICS LETTERS, vol. 26, no. 24, 22 November 1990, STEVENAGE, HERTS., GB, pages 2020 - 2022 VUCETIC B. & ZHANG L.: 'CONSTRUCTION OF BLOCK MODULATION CODES OVER RINGS FOR FADING CHANNELS' * page 2021, right column, line 24 - line 38 *	1	

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The present search report has been drawn up for all claims

Place of search	Date of completion of the search	Examiner
THE HAGUE	26 APRIL 1993	GHIGLIOTTI L.

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<tr><td colspan="2">European Patent Office</td><td>EUROPEAN SEARCH REPORT</td><td>Application Number
EP 92 31 0582
Page 2</td></tr>
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			TECHNICAL FIELDS SEARCHED (Int. Cl.5)

The present search report has been drawn up for all claims

Place of search	Date of completion of the search	Examiner
THE HAGUE	26 APRIL 1993	GHIGLIOTTI L.

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